PURDUE UNIVERSITY
SCHOOL OF ELECTRICAL ENGINEERING
ELECTRONIC SYSTEMS RESEARCH LABORATORY

SEMI-ANNUAL REPORT OF RESEARCH

PERFORMED UNDER GRANT NsG–553

January 1, 1967 through June 30, 1967

LAFAYETTE, INDIANA

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Foreword

This report summarizes work carried out at the Electronic Systems Research Laboratory of Purdue University under NASA Grant NsG-553 during the period January 1, 1967 through June 30, 1967.

In keeping with NASA's policy for administration of research grants, the report has been kept as concise as possible and, when appropriate, reference has been made to interim reports, internal memoranda, and technical papers resulting from research carried out under this grant.

The format of the report consists of a listing of technical papers and internal memoranda which have been submitted to NASA during the period covered by the report, abstracts of interim reports submitted during the reporting period, and appropriate extracts from the current Purdue University, School of Electrical Engineering, Semi-Annual Research Summary.

John C Lindenlaub

John C. Lindenlaub,
Principal Investigator

I. Technical Papers and Internal Memoranda

Two technical papers were presented at the 1967 IEEE International Conference on Communication held in Minneapolis, Minnesota, June 12-14, 1967.

i) "Performance of Self Bit Syncronization Systems" by E. J. Luecke and P. A. Wintz.

ii) "Effects of a Class of Phase Comparators on the Threshold and Lock Range of Phase Lock Loop Systems," by J. J. Uhran and J. C. Lindenlaub.

Copies of the first paper were sent to NASA Headquarters on June 26, 1967. The second paper was based on results contained in the interim report "Threshold Studies of Phase Lock Loop Systems" published in December 1966 and summarized in our last semi-annual report.

Electronic Systems Research Laboratory Memorandum Report 67-2, "Equivalence of Phase Lock Loop Systems and a Discriminator followed by a Nonlinear Feedback Filter," by J. C. Lindenlaub and D. P. Olsen, is in the process of being duplicated, and copies will be sent to NASA Headquarters and NASA Center reviewers as they become available.

II. Abstracts of Interim Reports

An interim report entitled "A New Class of Cyclic Codes" by D. R. Anderson has been completed and copies of the report mailed to NASA Headquarters on June 26, 1967. The report has also been accepted for publication in the SIAM Journal. An abstract of this report is given below:

The purpose of this work is to consider a new class of error-correcting codes, which shall be called primitive root codes (PR-codes). The class of PR-codes is made up of cyclic codes and forms a generalization of the Reed-Solomon codes. However, PR-codes differ significantly from BCH-codes in that each PR-code is an intersection (usually non-trivial) of BCH-codes. Thus our approach to code generation is different from that of Solomon and Mattson, who obtain the BCH-codes as the natural generalization of the Reed-Solomon codes.

Our results on PR-codes can be summarized as follows. It is first shown that for many lengths they are more numerous than BCH-codes, so that PR-codes are not just BCH-codes in another guise. Certain simple results on minimum distance and error-correcting properties are obtained, and in the process each PR-code is shown to be the intersection of a number of BCH-codes. For a large sub-class of the PR-codes in which all possible code lengths are represented asymptotic upper and lower estimates of the minimum distance are obtained. This result turns out to yield upper estimates of the minimum distance of many low information rate BCH-codes. Finally, the family of equal-period linear recurring (linear shift register) sequences corresponding to each of the latter codes are shown to have two striking properties. On the one hand, every sequence is a linear combination

over GF(p) of translates of equal-period maximum length linear recurrent sequences with values in GF(p), and, on the other hand, in the binary case any two sequences in a given family have a pseudo-random property relative to each other; namely, the number of agreements less the number of disagreements between a period of one and any translate of a period of the other is of the order of the square root of the period.

III. Research Summaries

A. A Receiver System for Digital Communication over Frequency Selective Channels

J. C. Lindenlaub

C. C. Bailey

The performance of an adaptive receiver system for digital communication over time-variant frequency selective channels is being investigated. The receiver scheme is based on an idea introduced by Sunde[1] in a study of probability of error performance of communication systems operating over "gaussian" random channels. Sunde showed that when the (instantaneous) filtering action of the channel is divided into its amplitude and phase components, the distortion introduced by the phase component will have far more effect on the communication signal than will the amplitude component. Furthermore, he noted that if a power series expansion of the channel's phase response function is made, the third term (involving the second derivative) of this series can be considered to have the most influence on the distortion introduced into the communication signal. The distortion attributed to this term in the expansion is called linear delay distortion. It is proposed that a useful receiver system for use with such random channels would be one which adaptively corrects for the linear delay distortion introduced by the channel. Such a receiver system must consist of a system to measure the channel's instantaneous linear delay distortion and a means of correcting the effect of this distortion. Measurement of the channel's linear delay distortion can be accomplished with the use of two transmitted pilot tones, situated just above and below the data transmission band. At the receiver, phase detectors can be used at each of the pilot tone frequencies to obtain a measurement of the linear delay distortion of the channel. A linear filter with unit amplitude and quadratic phase characteristics can be used for correction of the phase distortion.

Several aspects of the performance of this system are of interest. The first of these is how well such a system will perform under ideal conditions, i.e., with perfect measurement of the amount of linear delay distortion present and perfect correction of this distortion. It is also of interest to determine the effect of the pilot tone measurement system on the receiver performance, i.e., how much degradation results from the error due to pilot-tone measurement. Finally, the question of the optimal division of power between pilot tones and data signals is being investigated. Each of these performance investigations is being carried out by means of digital computer simulation.

Reference

1. Sunde, E. D., "Digital Troposcatter and Modulation Theory," Bell System Technical Journal, Vol. 43, pp. 143-214; Jan., 1964, (Part I).

B. Analog Transmission over Dispersive Communications Channels

J. C. Lindenlaub

D. P. Murray

A signal passing through a dispersive transmission medium suffers distortion due to multipath and selective fading, as well as distortion due to additive thermal or shot noise. In an effort to improve the performance of communications systems operating over dispersive channels, research has been undertaken to find techniques for overcoming the effects of dispersion on analog signals, and to evaluate the performance of these techniques.

Before dispersion can be counteracted, it is first necessary to measure the state of the channel. These measurements can be made on the basis of the message-bearing signal alone, or by means of a separate transmitted-reference signal. The first technique leads to non-linear and unrealizable receiver structures, while the second requires the expending of part of the available transmitter power in channel measurement. However, the transmitted reference technique leads to simple receiver structures. It also gives insight into the nature of the channel. Research thus far has concentrated on transmitted-reference techniques of measuring the channel parameters. In order to do this in an efficient manner, a joint optimization of the transmitted reference signals and channel state estimators has been carried out; the optimal system and signals have been found, and performance evaluated.

Once the state of the channel is known reasonably well, it is possible to find physically realizable maximum likelihood (and sometimes minimum variance) estimators for analog messages sent over the channel.

These optimal estimation systems, or demodulators, have been found for A.M. and P.M. For the A.M. case performance, it has been evaluated for a variety of different channel states. Simple suboptimal techniques have also been investigated. It has been found that there is an irreducible loss in the performance of communications systems operating through dispersive media compared with non-dispersive media; the exact loss depends upon the channel state. The suboptimal techniques have performance which ranges from optimal to awful, depending upon the particular channel state.

At present the performance of the aforementioned schemes is being evaluated in the more realistic case where both channel estimates and message estimates are simultaneously in error. Strategies include both a pure transmitted-reference technique and a combination transmitted-reference-sequential-feedback technique. It is hoped to determine the best trade-off between putting power into the message and power into channel estimation.

C. Source Encoding of Weather Satellite Data

P. A. Wintz

J. E. Essman

The basic problem of interest here is the representation of analog signals in the so called "signal vector space." It is desired to represent the signal by a k dimension vector (denoted by $\underline{S}$) in some "optimum manner." For this investigation, the phrase "optimum manner" means that one is to determine a set of suitable basis functions in signal space so as to minimize k, the dimension of the signal vector $\underline{S}$, and still maintain an acceptable approximation error ϵ.

An analytical approach has been postulated where, out of the totality of functions which are square integrable, one selects two subclasses L' and L" by imposing certain constraining conditions on the functions. The constraints might be that the functions in L' satisfy some differential equation and the functions in L" be required to satisfy certain conditions on the spectrial density; i.e., $B^n = \frac{\int f^n S(f)\, df}{\int S(f)\, df}$. The problem then reduces to one of choosing a suitable set of basis functions in the intersection sub space L' L".

In view of the mathematical difficulties encountered in the above problem, an experimental approach is being conducted using pictorial data from the Nimbus II meteorological satellite. The receiving station utilizes a 30 foot parabolic antenna, a low noise nuvistor preamplifier and a 136.95 M Hz FM receiver with a 40 K Hz IF bandwidth. The output of the F.M. discriminator is a 2400 Hz signal on which the pictorial data is amplitude modulated. The necessary amplication, demodulators, and synchronization using phase lock loop techniques have been constructed in order to obtain useful data to be used in the experimental analog to digital conversion program.

Various sampling filters (weighting functions) and the associated reconstruction filters have been designed and set up on the analogue computer, and the data obtained from Nimbus II used as the input signal. The evaluation of the approximation is made by both mean-squared error measurements and by qualitatively comparing the pictures reproduced from the sampled signal and the actual signal. A comparison is also made with "conventional" sampling techniques. To date, first order and second order filters have been investigated. These include the use of the well-known Laquerre functions as the basis. Results indicate that for a particular type of a signal spectrum there is a best set of basis functions.

Future plans include an investigation of higher order filters, a possibility of an adaptive filter, and an increased effort in the solution of the analytial problem.

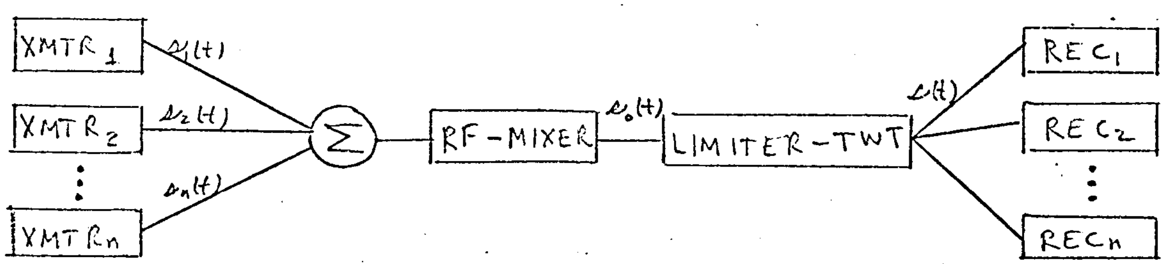
D. Analysis of a Multiple Access Satellite Communication System

D. R. Anderson

P. A. Wintz

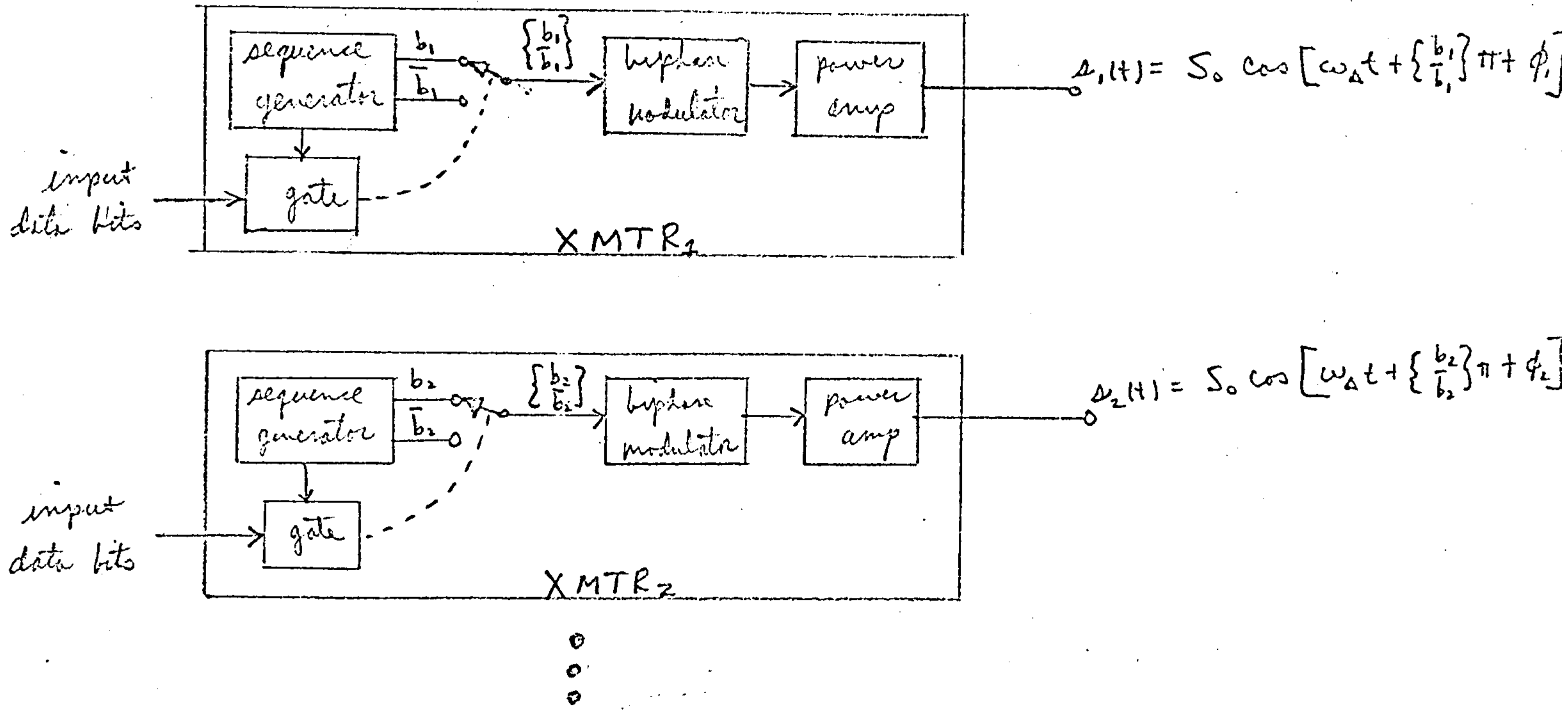
A model for a multiple access satellite communication system is presented in Fig. 1. The n active transmitters simultaneously transmit through the satellite repeater to the n receivers. Each transmitter contains a PN sequence generator, a biphase modulator, and a power amplifier as shown in Fig. 2. The PN sequence generator for transmitter #1 generates the PN sequence $b_1 = b_1(t)$ and its complement $\overline{b}_1 = -b_1(t)$. (A typical sequence is illustrated in Fig. 3). The input to the biphase modulator is a sequence of biphase signals, each one of which is either a period of $b_1(t)$ or a period of $\overline{b}_1(t)$. We label this sequence $\left\{ \frac{b_1}{\overline{b}_1} \right\}$. Either PSK or DPSK encoding can be used to encode the input data into the sequence $\left\{ \frac{b_1}{\overline{b}_1} \right\}$. The biphase modulator phase modulates a carrier with the sequence $\left\{ \frac{b_1}{\overline{b}_1} \right\}$.

The remaining n - 1 transmitters are identical to the first except that a different PN sequence is used for each transmitter, i.e., the ith transmitter uses the sequence $b_i = b_i(t)$ and its complement $\overline{b}_i = -b_i(t)$. As illustrated in Fig. 3, the n transmitters need not be in either information bit synchronization or in data bit synchronization. We do, however, assume that the data bit durations Δt and the information bit durations $T = (a^N-1)\Delta t$ are the same for all transmitters. Hence, all n sequences



XMTR 1
XMTR 2
XMTR n
$s_1(t)$
$s_2(t)$
$s_n(t)$
RF-MIXER
$s_0(t)$
LIMITER-TWT
$s(t)$
REC 1
REC 2
REC n

Fig. 1



sequence generator
b_1
$\bar{b}_1$
$\begin{Bmatrix} b_1 \\ \bar{b}_1 \end{Bmatrix}$
biphase modulator
power amp
gate
input data bits
XMTR 1
$s_1(t) = S_0 \cos\left[\omega_\Delta t + \begin{Bmatrix} b_1 \\ \bar{b}_1 \end{Bmatrix}\pi + \phi_1\right]$
sequence generator
b_2
$\bar{b}_2$
$\begin{Bmatrix} b_2 \\ \bar{b}_2 \end{Bmatrix}$
biphase modulator
power amp
gate
input data bits
XMTR 2
$s_2(t) = S_0 \cos\left[\omega_\Delta t + \begin{Bmatrix} b_2 \\ \bar{b}_2 \end{Bmatrix}\pi + \phi_2\right]$

Fig. 2

b_i, $i = 1,\ldots,n$ are the same length, and all transmitters operate at the same data rate. The particular set of n PN sequences, their properties, and methods for generating them are discussed in [1]. The signal at the output of the satellite repeater contains the n signal terms plus the so-called modulation products generated in the hard limiter.

The receiver for transmitter #1 is shown in Fig. 4. The signal appearing at the receiver input is identical to the repeater output except for an amplitude scale factor and a phase shift. Therefore, the signal at the output of the receiver RF-MIXER-IF stage is given by

$$x(t) = s(t) + n(t)$$

$n(t)$ represents the receiver noise for which we assume a power per unit bandwidth of N_o watts/Hz.

The correlation detector operates in a coherent mode. It is coherent both in carrier phase and in information bit synchronization. This is equivalent to assuming the availability of a reference signal of the form

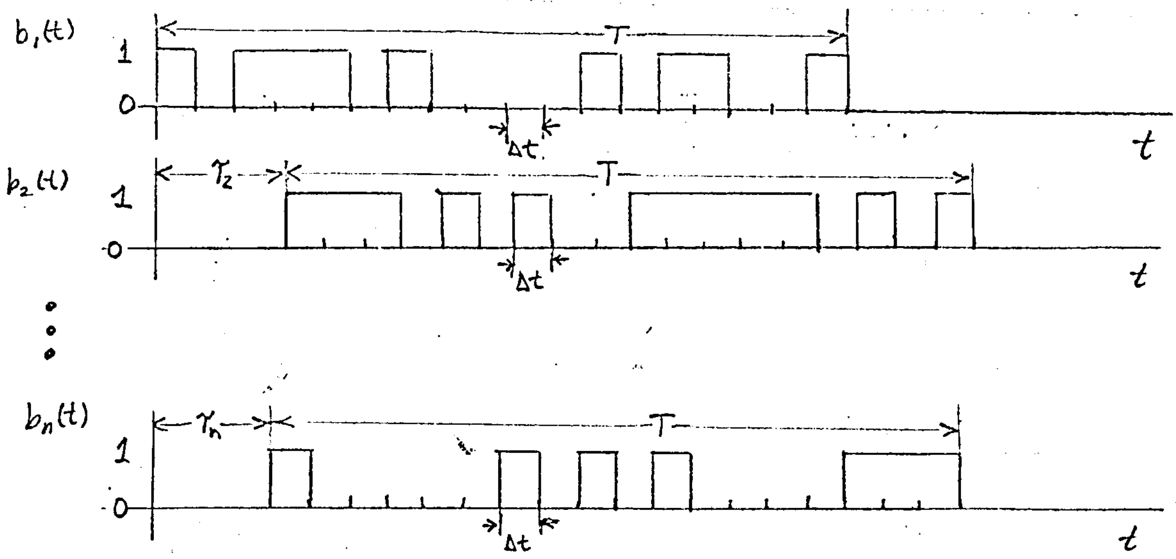
$$s_R(t) = \cos\left[\,\omega t + b_1\,\pi + \varphi_1\,\right] \tag{1}$$

that is both in phase and information bit synchronization with the signal term of $s(t)$ due to transmitter #1. Methods for generating such a coherent reference signal from the waveform $x(t)$ are discussed in [1]. The correlation detector computes the test static

$$\alpha = \frac{1}{T}\int_o^T x(t)\; s_R(t)\; dt \tag{2}$$

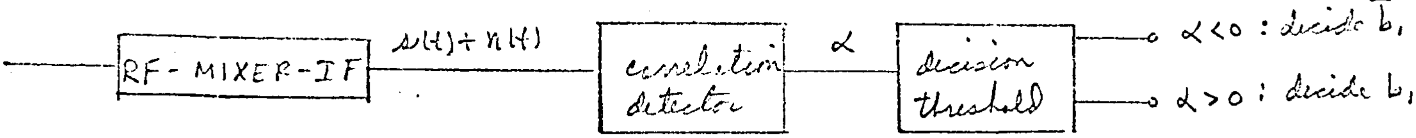
and announces a decision in accordance with the decision rule

$$\begin{aligned} &\alpha > 0: \quad \text{announce } b_1 \text{ transmitted} \\ &\alpha < 0: \quad \text{announce } \overline{b}_1 \text{ transmitted} \end{aligned} \tag{3}$$



$b_1(t)$
1
0
T
Δt
t
$b_2(t)$
τ_2
T
Δt
$b_n(t)$
τ_n
T
Δt

Fig. 3



RF - MIXER - IF
$s(t) + n(t)$
correlation detector
α
decision threshold
$\alpha < 0$: decide $\bar{b}_1$
$\alpha > 0$: decide b_1

Fig. 4

The probability of a detection error is the probability that the test statistic α is less than zero, i.e.,

$$P_E = \text{Prob}\left[\ \alpha < 0\ \right] = \int_{-\infty}^{0} p(\alpha)\ d\alpha \tag{4}$$

where $p(\alpha)$ is the probability density function for the random variable α. We can show $p(\alpha)$ is Gaussian using a precise form of the well-known emperical rule that narrowband-filtering of a wide-band process results in a Gaussian output. P_E depends on the detected SNR = $\bar{\alpha}^2/\overline{(\alpha-\bar{\alpha})^2}$. In [1] we have shown that

$$\text{SNR} = \frac{.88}{n\ \pi_c^2/4} \tag{5}$$

where π_c is the peak of the in and out of phase crosscorrelations of the n sequences. In [1] we show how to choose M PN sequences address such that

$$\pi_c^2 = \frac{4\ M^2}{2^N-1} \tag{6}$$

where M is the total number of transmitters (PN sequences). (Recall that n is the number of active transmitters.) Therefore,

$$\text{SNR} = (.88)\ \frac{2^N-1}{n\ M^2} \tag{7}$$

Suppose that all M transmitters are active and that all M transmitted signals use the entire repeater bandwidth. Then the data rate R is given by

$$R = \frac{1}{T} = \frac{1}{(2^N-1)\ \Delta t} = \frac{B}{2(2^N-1)} \tag{8}$$

where we have defined the bandwidth B as

$$B = \frac{2}{\Delta t} \tag{9}$$

Then using Eq. (7) we have

$$M = \left[\frac{(.44)}{(SNR)} \frac{B}{R} \right]^{1/3} \tag{10}$$

For example, for $R = 10^3$ information bits/sec/channel, $B = 10^8$ Hz, SNR = 6.6 (8 db) which gives $P_E = 10^{-4}$, we get M = 19 channels.

Considerably more channels can be accommodated if all of the transmitters do not use the total repeater bandwidth. Suppose each transmitted signal has bandwidth βHa, and that a guard band of γ Hz is used between channels. Then, using Eq. (10), we can accommodate

$$M' = \left[\frac{(.44)}{(SNR)} \frac{\beta}{R} \right]^{1/3}$$

channels in any band of βHz and $\frac{B}{\beta + \gamma}$ such bands in the total bandwidth of BHz, i.e.,

$$M = \frac{B}{\beta + \gamma} \left[\frac{(.44)}{(SNR)} \frac{\beta}{R} \right]^{1/3}$$

As an example we note that using a bandwidth of β = 9 MHz per channel, an information bit rate per channel of $R = 10^3$ bits, and SNR = 6.6 (P_E=10^{-4}) gives M' = 8. However, in a 100 MHz bandwidth 10 such bands can be accommodated if we allow 1 MHz guard bands. Therefore, a total of M = 80 channels is possible.

Reference

1. Anderson, D. R., and P. A. Wintz "Comments on Satellite Communications," Internal Memorandum, School of Electrical Engineering, Purdue University, Lafayette, Indiana; March, 1967. (Submitted for publication to IEEE Trans. on Communication Technology.)

E. A Phase Lock Loop System with Linear Modulo $2n\pi$ Phase Detector

J. C. Lindenlaub

D. P. Olsen

The construction of a phase lock loop having a phase detector with a linear modulo $2n\pi$ phase characteristic is complete. There are a few modifications of the previous system block diagram.[1] The new block diagram is shown in Fig. 1. Part of the testing of this system for proper operation is completed.

Since many experiments are planned which require an FM signal plus band limited Gaussian noise, a test fixture was designed and built to generate the necessary test signals. Fig. 2 is a block diagram of this device. With this test set one can conveniently and independently vary the modulation, signal, and noise levels as well as the signal center frequency.

A theoretical analysis of the threshold characteristics of the linear modulo $2n\pi$ phase lock system has been initiated.

<u>Reference</u>

1. "A Phase Lock Loop System with Modulo $2n\pi$ Phase Detector," by J. C. Lindenlaub and D. P. Olsen, Purdue University, School of Electrical Engineering, 5th Semi-Annual Research Summary, p. 156; December, 1966.

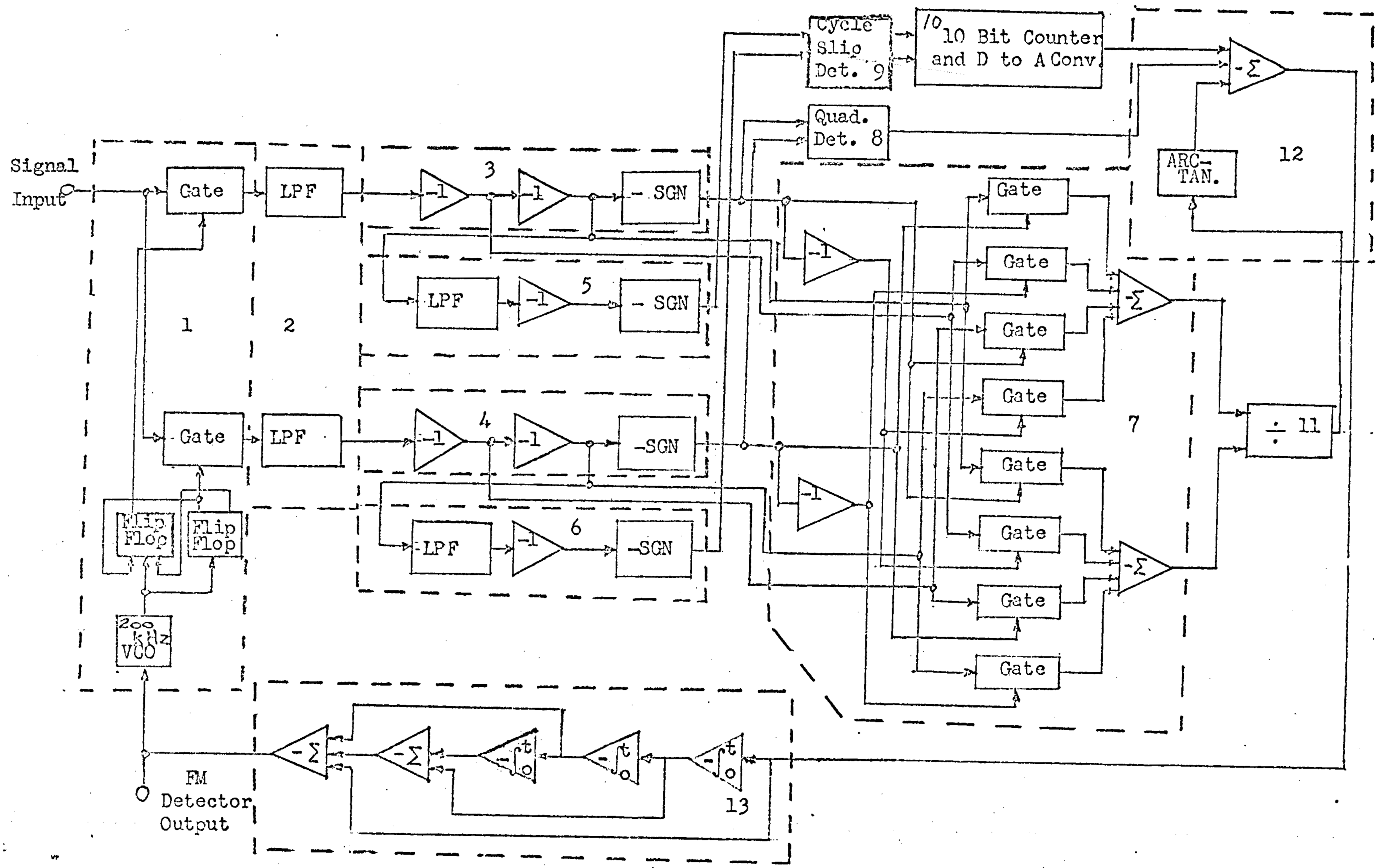


Fig. 1. Linear Modulo 2 nπ Phase Lock Loop

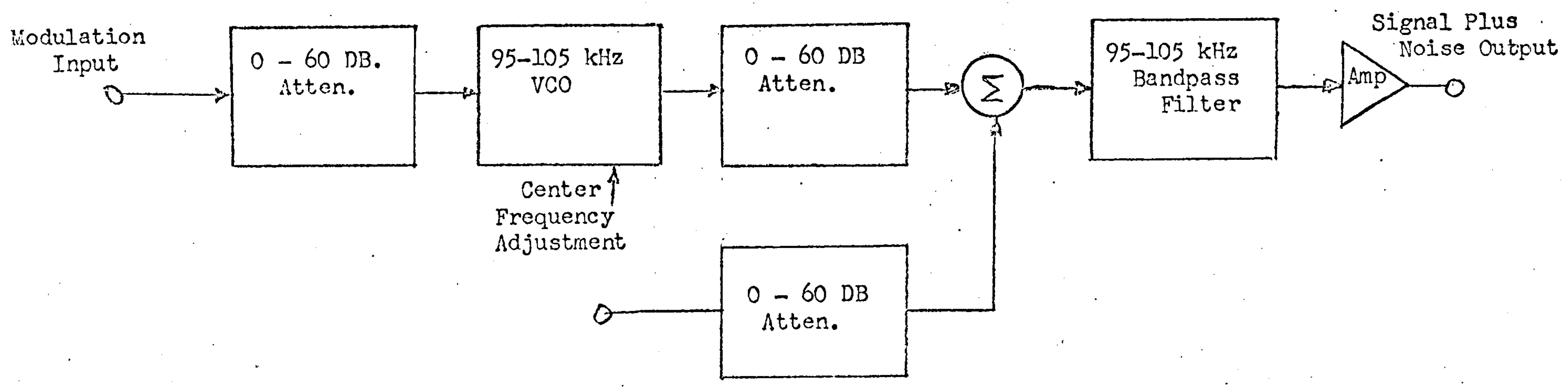


Fig 2. Noise and FM Signal Generator